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13 Linear Control Design and Simulation of Power System Stability with FACTS

Inter-area oscillations in power systems are triggered by, for example, disturbances such as variation in load demand or the action of voltage regulators due to a short circuit. The primary function of the damping controllers is to minimize the impact of these disturbances on the system within the limited dynamic rating of the actuator devices (excitation systems, FACTS-devices). In H_∞ control term, this is equivalent to designing a controller that minimizes the infinity norm of a chosen mix of closed-loop quantities.

The concept of H_∞ techniques for power system damping control design is about ten years old [1]-[5]. An interesting comparison between various techniques is made in [6]. There are two approaches for solving a standard H_∞ optimization problem: analytical and numerical. While the analytical approach seeks a positive semi definite solution to the Riccati equation [7], the numerical approach is to solve the Riccati inequality to optimize the relevant performance index. Although the Riccati inequality is non-linear, there are linearization techniques to convert it into a set of linear matrix inequalities (LMIs) [8][7], which simplifies the computational process.

The analytical approach is relatively straightforward but generally produces a controller that suffers from pole-zero cancellations between the plant and the controller [9]. The closed-loop damping ratio, which is very important in power system control design, can not be captured in a straight forward manner in a Riccati based design [10]. The numerical approach to the solution, using the linear matrix inequality (LMI) approach, has a distinct advantage as these design specifications can be addressed as additional constraints. Moreover, the controllers obtained using a numerical approach do not, in general, suffer from the problem of pole-zero cancellation [11].

Application of the H_∞ approach using LMIs has been reported in [12][13] for design of power system stabilizers (PSS). A mixed-sensitivity approach with an LMI based solution was applied for the design of damping control for superconducting magnetic energy storage (SMES) devices [10][14][15]. Recently this approach has been extended for the design of damping control provided by different FACTS-devices [16][17][5]. This chapter describes the basic concept of mixed-sensitivity design formulation with the problem translated into a generalized H_∞ problem [8][7]. The entire control design methodology is illustrated by a couple of case studies on a study power system model. The damping control performance is validated in both frequency domain and time domain.

The second half of this chapter focuses on extending these design techniques to a time delayed system. We assume that in centralized design remote signals are instantly available. However, in reality, depending on signal transmission protocol, a delay is introduced. This would transform the system into a delayed system, which the control algorithm must take into consideration. We have applied a predictor based approach [5]. An SVC is used to damp oscillations through a delayed remote signal. The performance of the control has been validated on the same study system and conclusions are made.

13.1 H-Infinity Mixed-Sensitivity Formulation

The standard mixed-sensitivity formulation for output disturbance rejection and control effort optimization is shown in Fig. 13.1, where $G(s)$ is the open loop system model and $K(s)$ is the controller to be designed. The sensitivity $S = (I - GK)^{-1}$ represents the transfer function between the disturbance input $w(s)$ and the measured output $y(s)$. In the case of a power system, typically the sensitivity S is the impact of load changes on the oscillations of angular or machine speed. So it is required to minimize $\|S\|_{\infty}$. It is also required to minimize H_{∞} norm of the transfer function between the disturbance and the control output to optimize the control effort within a limited bandwidth. This is equivalent to minimizing $\|KS\|_{\infty}$. Thus, the minimization problem can be summarized as follows:

$$\min_{K \in S} \left\| \begin{bmatrix} S \\ KS \end{bmatrix} \right\|_{\infty} \quad (13.1)$$

where S is the set of all internally stabilizing controllers K .

It is, however, not possible to simultaneously minimize both S and KS over the whole frequency spectrum. This is not required in practice either. The disturbance rejection is usually required at low frequencies. Thus S can be minimized over the low frequency range where as, KS can be minimized at higher frequencies where limited control action is required.

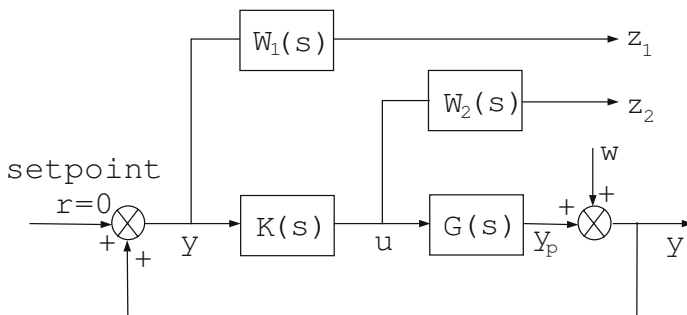


Fig. 13.1. Mixed-sensitivity formulation

Appropriate weighting filters $W_1(s)$ and $W_2(s)$ are used to emphasize the minimization of each individual transfer function at the different frequency ranges of interest. The minimization problem is formulated such that S is less than $W_1(s)^{-1}$ and KS is less than $W_2(s)^{-1}$. The standard practice, therefore, is to select $W_1(s)$ as an appropriate low pass filter for output disturbance rejection and to select $W_2(s)$ as a high-pass filter to reduce the control effort over the high frequency range. The problem can be restated as follows:

find a stabilizing controller, such that:

$$\min_{K \in S} \left\| \begin{bmatrix} W_1 S \\ W_2 K S \end{bmatrix} \right\|_{\infty} < 1 \quad (13.2)$$

13.2 Generalized H-Infinity Problem with Pole Placement

The mixed-sensitivity design problem is translated into a generalized H_{∞} problem. The first step is to set up a generalized regulator P corresponding to the mixed-sensitivity formulation. For simplicity, it is assumed that the weights W_1 and W_2 are not present but will be taken care of later. Without the weights, the mixed-sensitivity formulation in Fig. 13.1 can be redrawn in terms of the A , B and C matrices of the system, as shown in Fig. 13.2. Without any loss of generality, it can be assumed that $D=0$.

From Fig. 13.2, it can be readily seen that:

$$\dot{x} = Ax + Bu \quad (13.3)$$

$$z_1 = Cx + w \quad (13.4)$$

$$z_2 = u \quad (13.5)$$

$$y = Cx + w \quad (13.6)$$

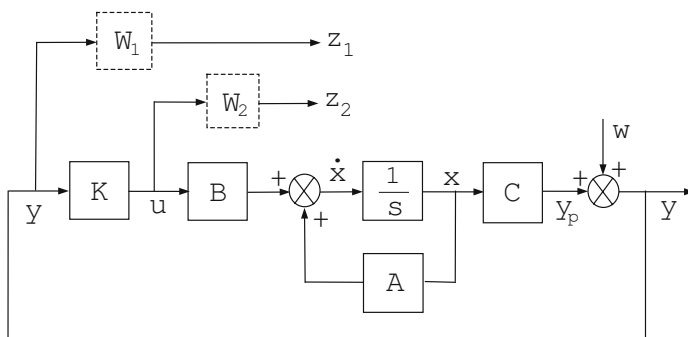


Fig. 13.2. Generalized regulator set-up for mixed-sensitivity formulation

The state-space representation of a generalized regulator P is given as:

$$\begin{bmatrix} \dot{x} \\ z_1 \\ z_2 \\ y \end{bmatrix} = \begin{bmatrix} A & 0 & B \\ C & I & 0 \\ 0 & 0 & I \\ C & I & 0 \end{bmatrix} \begin{bmatrix} x \\ w \\ u \end{bmatrix} \quad (13.7)$$

- x : state variable vector of the power system (e.g. machine angle, machine speed etc),
 w : disturbance input (e.g. a step change in excitation system reference),
 u : control input (e.g. output of PSS or FACTS-devices),
 y : measured output (e.g. power flow, line current, bus voltage etc),
 z : regulated output.

For the weighting filters of the generalized regulator, i.e. the state-space representations of W_1 and W_2 , are placed in a diagonal form using the *sdiag* function available in Matlab [18]. The result is multiplied with P (without the weights) using the *smult* function also available in Matlab.

The task now is to find an LTI control law $u = Ky$ for some H_∞ performance index $\gamma > 0$, such that $\|T_{wz}\|_\infty < \gamma$ where, T_{wz} denotes the closed-loop transfer function from w to z . If the state-space representation of the LTI controller is given by:

$$\begin{aligned} \dot{x}_k &= A_k x_k + B_k y \\ u &= C_k x_k + D_k y \end{aligned} \quad (13.8)$$

then the closed-loop transfer function T_{wz} from w to z is given by $T_{wz}(s) = D_{cl} + C_{cl}(sI - A_{cl})^{-1}B_{cl}$ where,

$$A_{cl} = \begin{bmatrix} A + B_2 D_k C_2 & B_2 C_k \\ B_k C_2 & A_k \end{bmatrix} \quad (13.9)$$

$$B_{cl} = \begin{bmatrix} B_1 + B_2 D_k D_{21} \\ B_k D_{21} \end{bmatrix} \quad (13.10)$$

$$C_{cl} = [C_1 + D_{12} D_k C_2 \quad D_{12} C_k] \quad (13.11)$$

$$D_{cl} = D_{11} + D_{12} D_k D_{21} \quad (13.12)$$

In addition to guaranteeing robustness by satisfying $\|T_{wz}\|_\infty < \gamma$, another design requirement in power systems is to ensure that the oscillations settle within 10-15 s [14]. This is achieved if the closed-loop poles corresponding to the critical modes have the minimum damping ratio. In consideration of this, the above problem statement can be modified to include the pole-placement constraint so that the problem is now: *find an LTI control law $u = Ky$ such that:*

- $\|T_{wz}\|_\infty < \gamma$
- Poles of the closed-loop system lie in D

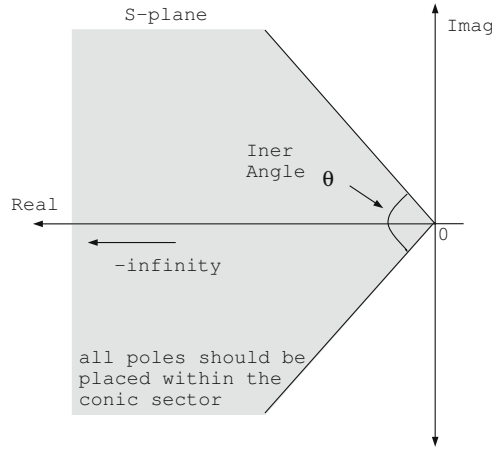


Fig. 13.3. Conic sector region for pole-placement

D defines a region in the complex plane having certain geometric shapes like disks, conic sectors, vertical/horizontal strips, etc. or intersections of these. A ‘conic sector’, with inner angle θ and apex at the origin is an appropriate region for power system applications as it ensures a minimum damping ratio

$$\zeta_{\min} = \cos^{-1} \frac{\theta}{2} \text{ for the closed-loop poles.}$$

13.3 Matrix Inequality Formulation

The bounded real lemma [11] and Schur’s formula for the determinant of a partitioned matrix [7], enable one to conclude that the H_∞ constraint $\|T_{wz}\|_\infty < \gamma$ is equivalent to the existence of a solution $X_\infty = X_\infty^T > 0$ to the following matrix inequality:

$$\begin{pmatrix} X_\infty A_{cl} + A_{cl}^T X_\infty & B_{cl} & X_\infty C_{cl}^T \\ B_{cl}^T & -\gamma I & D_{cl}^T \\ C_{cl} X_\infty & D_{cl} & -\gamma I \end{pmatrix} < 0 \quad (13.13)$$

A ‘conic sector’ with inner angle θ and apex at the origin is chosen as the region D within which the pole-placements are confined to. The closed-loop system matrix A_{cl} has all its poles inside the conical sector D if and only if there exists $X_D = X_D^T > 0$, such that the following matrix inequality is satisfied [21].

$$\begin{pmatrix} \sin \theta (A_{cl} X_D + X_D A_{cl}^T) & \cos \theta (A_{cl} X_D - X_D A_{cl}^T) \\ \cos \theta (X_D A_{cl}^T - A_{cl} X_D) & \sin \theta (X_D A_{cl}^T + A_{cl} X_D) \end{pmatrix} < 0 \quad (13.14)$$

The design specifications are feasible if and only if (13.13) and (13.14) hold for some positive semi-definite matrices X_∞ and X_D and some controller K with state-space $(A_k, B_k, C_k, \text{ and } D_k)$. However, the problem is not jointly convex in X_∞ and X_D unless it is solved for the same matrix X . In view of this, the sub-optimal H_∞ problem with pole-placement can be stated as follows:

find $X > 0$ and a controller K , such that (13.13) and (13.14) are satisfied with $X = X_\infty = X_D$ [20][21].

The inequalities (13.13) and (13.14) containing $A_{cl}X$ and $C_{cl}X$ are functions of the controller parameters, which themselves are functions of X . This makes the products $A_{cl}X$ and $C_{cl}X$ non-linear in X . However, a change of controller variables can convert the problem into a linear one. This will be described in the next section.

13.4 Linearization of Matrix Inequalities

The controller variables are implicitly defined in terms of the (unknown) matrix X . Let X and X^{-1} be partitioned as:

$$X = \begin{pmatrix} R & M \\ M^T & U \end{pmatrix}, X^{-1} = \begin{pmatrix} S & N \\ N^T & V \end{pmatrix} \quad (13.15)$$

For $\Pi_1 = \begin{pmatrix} R & I \\ M^T & 0 \end{pmatrix}$ and $\Pi_2 = \begin{pmatrix} I & S \\ 0 & N^T \end{pmatrix}$, X satisfies the identity $X\Pi_1 = \Pi_2$. The new controller variables are defined in (13.16) to (13.19).

$$\hat{A} = NA_kM^T + NB_kC_2R + SB_2C_kM^T + S(A + B_2D_kC_2)R \quad (13.16)$$

$$\hat{B} = NB_k + SB_2D_k \quad (13.17)$$

$$\hat{C} = C_kM^T + D_kC_2R \quad (13.18)$$

$$\hat{D} = D_k \quad (13.19)$$

The identity $XX^{-1} = I$ together with (13.15) gives:

$$MN^T = I - RS \quad (13.20)$$

If M and N have full row rank, then the controller matrices $A_k, B_k, C_k, \text{ and } D_k$ can always be computed from $\hat{A}, \hat{B}, \hat{C}, \hat{D}, R, S, M$ and N . Moreover, the controller matrices can be determined uniquely if the controller order is chosen to be equal to that of the generalized regulator [21].

Pre- and post-multiplying the inequality $X > 0$ by Π_2^T and Π_2 respectively, and carrying out appropriate change of variables according to (13.16), (13.17), (13.18) and (13.19) allows obtaining the following linear matrix inequality (LMI):

$$\begin{pmatrix} R & I \\ I & S \end{pmatrix} > 0 \quad (13.21)$$

Similarly, by pre- and post-multiplying the inequality (13.13) by $\text{diag}(\Pi_2^T, I, I)$ and $\text{diag}(\Pi_2, I, I)$ respectively and carrying out appropriate change of variables according to (13.16), (13.17), (13.18) and (13.19), the following LMI is obtained.

$$\begin{bmatrix} \Psi_{11} & \Psi_{21}^T \\ \Psi_{21} & \Psi_{22} \end{bmatrix} < 0 \quad (13.22)$$

where

$$\Psi_{11} = \begin{bmatrix} AR + RA^T + B_2 \hat{C} + C^T \hat{B}_2^T & B_1 + B_2 D \hat{D}_{21} \\ (B_1 + B_2 \hat{D} D_{21})^T & -\gamma I \end{bmatrix} \quad (13.23)$$

$$\Psi_{21} = \begin{bmatrix} \hat{A} + (A + B_2 \hat{D} C_2)^T & SB_1 + \hat{B} D_{21} \\ C_1 R + D_{12} \hat{C} & D_{11} + D_{12} \hat{D} D_{21} \end{bmatrix} \quad (13.24)$$

$$\Psi_{22} = \begin{bmatrix} A^T S + SA + \hat{B} C_2 + C_2^T \hat{B} & (C_1 + D_{12} \hat{D} C_2)^T \\ C_1 + D_{12} \hat{D} C_2 & -\gamma I \end{bmatrix} \quad (13.25)$$

Proceeding in a similar fashion by pre- and post-multiplying the inequality (13.14) by Π_2^T and Π_2 respectively, and carrying out the change of variables according to (13.16), (13.17), (13.18) and (13.19), the following LMI is obtained. A detailed explanation of this process can be found in [20][21].

$$\begin{pmatrix} \sin \theta(\phi + \phi^T) & \cos \theta(\phi - \phi^T) \\ \cos \theta(\phi^T - \phi) & \sin \theta(\phi^T + \phi) \end{pmatrix} < 0 \quad (13.26)$$

where

$$\phi = \begin{pmatrix} AR + B_2 \hat{C} & A + B_2 D \hat{D}_{21} \\ \hat{A} & SA + \hat{B} C_2 \end{pmatrix} \quad (13.27)$$

The system of LMIs in (13.21), (13.22) and (13.26) are solved for $R, S, \hat{A}, \hat{B}, \hat{C}$ and $\hat{D}$. A full-rank factorization of the matrix $I-RS$ is computed via singular value decomposition (SVD) such that $MN^T = I-RS$ holds for M and N being square and invertible. With known values of $R, S, \hat{A}, \hat{B}, \hat{C}, \hat{D}, M$ and N , the system of linear equations (13.16), (13.17), (13.18) and (13.19) can be solved for D_k, B_k, C_k and A_k in that order. The controller K is obtained and the resultant controller places the closed-loop poles in D and satisfies $\|T_{wz}\|_\infty < \gamma$.

13.5 Case Study

In this section, the prototype power system model, described in chapter 12, is used to illustrate the control design methodology in detail. The performance and robustness of the design is also validated.

13.5.1 Weight Selection

As mentioned earlier, the standard practice in H_∞ mixed-sensitivity design is to choose the weight $W_1(s)$ as an appropriate low pass filter for output disturbance rejection and choose $W_2(s)$ as a high-pass filter to reduce the control effort in the high frequency range. For the prototype system, the weights are accordingly chosen as follows:

$$\begin{aligned} W_1(s) &= \frac{30}{s + 30} \\ W_2(s) &= \frac{10s}{s + 100} \end{aligned} \quad (13.28)$$

The frequency responses of these weight functions are shown in Fig. 13.4. It can be seen that the two weights intersect at around 10 rad/s, noting that the critical modes to be controlled are below the frequency of 10 rad/s. Thus, the minimization of the sensitivity is emphasized up to this frequency and the control is constrained soon after.

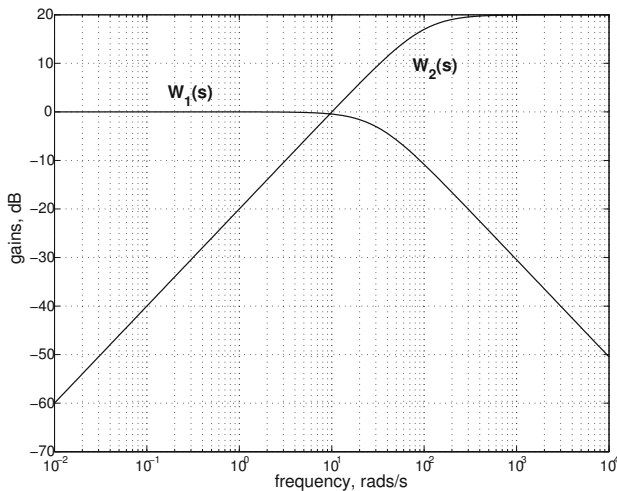


Fig. 13.4. Frequency response of weighting filter

13.5.2 Control Design

To facilitate the control design, and to reduce the complexity of the designed controller, the nominal system model was reduced to a 7th order equivalent as described in chapter 12. The generalized regulator problem was formulated according to (13.7) using the simplified system model and the weights in (13.28). The control design problem is to minimize γ such that (13.21), (13.22) and (13.26) are satisfied.

A series of functions which, are available with the *LMI toolbox* [22] in *Matlab* [18], is used to formulate and solve the optimization problem. The first step is to define the solution variables (also called the LMI variables). The variables $R, S, \hat{A}, \hat{B}, \hat{C}$ are defined using the appropriate Matlab toolbox function. The size of the variables and their structure is specified through this function. Having defined the solution variables, the next step is to set up the LMIs (13.21), (13.22) and (13.26) in terms of these variables. Each of the terms of an LMI and their respective positions are specified using the function from LMI toolbox. In this design, a ‘conic sector’ of inner angle $2 \cos^{-1} 0.15$ with apex at the origin was chosen as the pole-placement region to ensure a minimum damping ratio of 0.15 for the closed-loop system. To achieve this, the value of θ in (13.26) was set to $\cos^{-1} 0.15$. The three sets of LMIs are combined in a system of LMI. The optimal values of $R, S, \hat{A}, \hat{B}, \hat{C}$ are retrieved from the output of the optimization function by using the suitable function from the toolbox. The control variables A_k, B_k, C_k and D_k are computed accordingly with the help of equations (13.16), (13.17), (13.18) and (13.19).

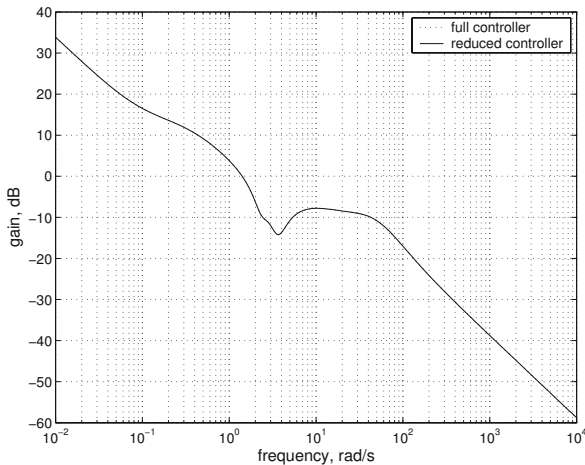


Fig. 13.5. Frequency response of full and reduced order controller

The order of the controller obtained from this design routine is equal to the order of the reduced system order plus the order of the weights. As there are three weights associated with the three measured outputs and one with the control input, the size of the designed controller is 14 (9+3+1). The designed controller was simplified further to a 10th order equivalent without affecting the frequency response, as shown in Fig. 13.5. The frequency response of the sensitivity S and the controller sensitivity product KS are plotted in Figs. 13.6 and 13.7.

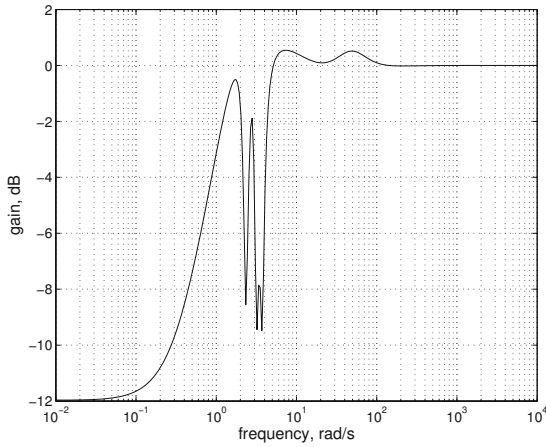


Fig. 13.6. Frequency response of sensitivity (S)

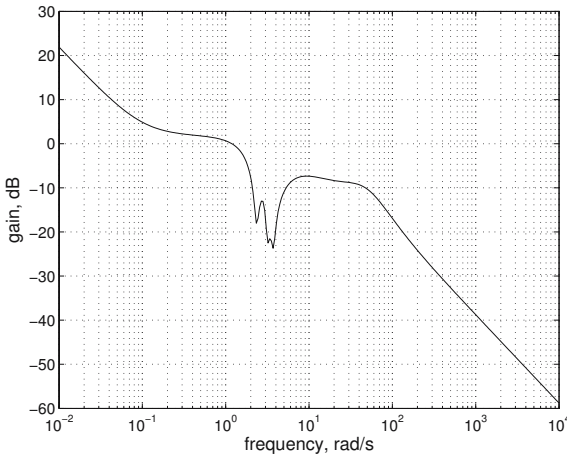


Fig. 13.7. Frequency response of control time sensitivity (KS)

As discussed before, S should be low at the lower frequencies to achieve disturbance rejection but comparatively high values can be tolerated at higher frequencies. This is achieved in the designed controller as seen from Fig. 13.6. In contrast, to ensure satisfactory performance, KS should be low at high frequencies to reduce the control effort.

The design steps can be summarized as follows:

- Simplify the system model;
- Formulate the generalized regulator using the simplified system model and the mixed-sensitivity weights;
- Define the LMI variables using the *lmivar* function;
- Construct the terms of the LMIs using the *lmiterm* function;
- Assemble the individual LMIs into a set of LMIs employing the *getlmis* function;
- Solve the γ optimization problem with the set of LMI constraints using the *mincx* function;
- Retrieve the optimum value of the solution variables through the *dec2mat* function;
- Determine the controller using the optimum value of the solution variables; and
- Simplify the designed controller.

Alternatively, the design problem can be solved by suitably defining the objectives in the argument of the function *hinfnix*, available with the *LMI Toolbox* [22] for Matlab [18]. The pole-placement constraint can be imposed by using the *lmireg* function, which is an interactive interface for specifying different LMI regions.

Table 13.1. Damping ratios and frequencies of the inter-area modes

Mode No.	Without Control		With Control	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
1	0.0626	0.3913	0.2336	0.3590
2	0.0435	0.5080	0.1316	0.5094
3	0.0554	0.6232	0.1456	0.6384
4	0.0499	0.7915	0.0550	0.7843

13.5.3 Performance Evaluation

The eigenvalues of the closed-loop system were computed to examine the performance of the designed controller in terms of improving the damping ratios of the inter-area modes. The results are summarized in Tables 13.1. It can be seen that the damping ratios of the three critical inter-area modes, shown in boldface, are improved in the closed loop.

It is to be noted that by imposing the pole-placement constraint, as described earlier, a minimum damping ratio of 0.15 could be ensured for the simplified closed-loop system. However, the results shown here are based on the full and

original system model. Therefore the damping ratios under certain situations are less than 0.15. Nonetheless, they are still adequate enough to ensure that oscillations settle within 12-15 s.

The damping action of the designed controller was examined under different types of disturbances in the system. These included, amongst others, changes in power flow levels over key transmission corridors and change in type of loads. Table 13.2 displays the damping ratios of the inter-area modes for a range of power flows across the interconnection between the areas NETS and NYPS in the study system.

The performance of the controller was tested with various load models including a constant impedance (CI), a mixture of constant current and constant impedance (CC+CI), a mixture of constant power and constant impedance (CP+CI) and with dynamic load characteristics. The damping ratios of the inter-area modes are listed in Table 13.3 for different types of load characteristics. From the damping ratios displayed in the tables below it can be concluded that the action of the designed controller is robust against widely varying operating conditions.

Table 13.2. Damping ratios and frequencies of the critical inter-area modes at different levels of power flow between NETS and NYPS

Power Flow (MW)	Mode 1		Mode 2		Mode 3	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
100	0.2420	0.3566	0.1374	0.5106	0.1351	0.6640
500	0.2371	0.3578	0.1338	0.5097	0.1419	0.6451
700	0.2336	0.3590	0.1316	0.5094	0.1456	0.6384
900	0.2300	0.3609	0.1289	0.5093	0.1491	0.6251

Table 13.3. Damping ratios and frequencies of the critical inter-area modes for different load models

Type of Load	Mode 1		Mode 2		Mode 3	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
CI	0.2336	0.3590	0.1316	0.5094	0.1456	0.6384
CI+CC	0.2313	0.3608	0.1308	0.5175	0.1314	0.6353
CI+CP	0.2251	0.3621	0.1309	0.5260	0.1175	0.6351
Dynamic	0.2304	0.3582	0.1399	0.5135	0.1456	0.6381

13.5.4 Simulation Results

One of the most severe disturbances, in terms of producing poorly damped inter-area oscillations, is a three-phase fault in one of the key transmission circuits. For temporary faults, the circuit breaker re-closes after few cycles and normal operation is restored. For a permanent fault, the line is tripped out of operation. The other types of disturbances in the system, such as change of load characteristics

and sudden change in power flow, are less severe compared to three faults and are not considered here.

The simulations were carried out to determine system performance during probable fault scenarios in the NETS and NYPS inter-connection. There are three inter-connections between NETS and NYPS connecting buses #60-#61, #53-#54 and #27-#53, respectively. Each of these inter-connections consists of two lines and an outage of one of these lines weakens the interconnection considerably. To examine the effect of such disturbances, a series of solid three-phase solid faults, each of about 80 ms (about 5 cycles) in duration, were simulated in the following locations:

- (a) bus #60 followed by auto-reclosing of the circuit breaker
- (b) bus #53 followed by outage of one of the tie-lines between buses #53-#54
- (c) bus #53 followed by outage of one of the tie-lines between buses #27-#53
- (d) bus #60 followed by outage of one of the tie-lines between buses #60-#61

Simulations were carried out in Matlab *Simulink* [23] for 25 s employing the *trapezoidal integration* method with a variable step size. The disturbance was created 1 s after the start of the simulation. The dynamic response of the system following the disturbance is shown in Figs. 13.8, 13.9 and 13.10. These figures illustrate the relative angular separation between the generators located in separate geographical regions. Inter-area oscillations are mostly manifested in these angular differences and are, therefore, chosen for display. It can be seen that inter-area oscillations settle within the desired performance specification of 12–15 s for a range of post-fault operating conditions. The TCSC was limited to provide between 0.1 to 0.8 p.u. of compensation. The variation in the percentage compensation provided by the TCSC is shown in Fig. 13.10.

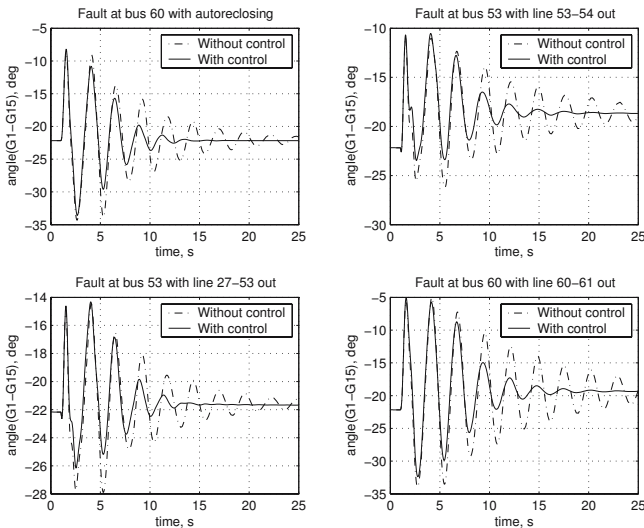


Fig. 13.8. Dynamic response of the system

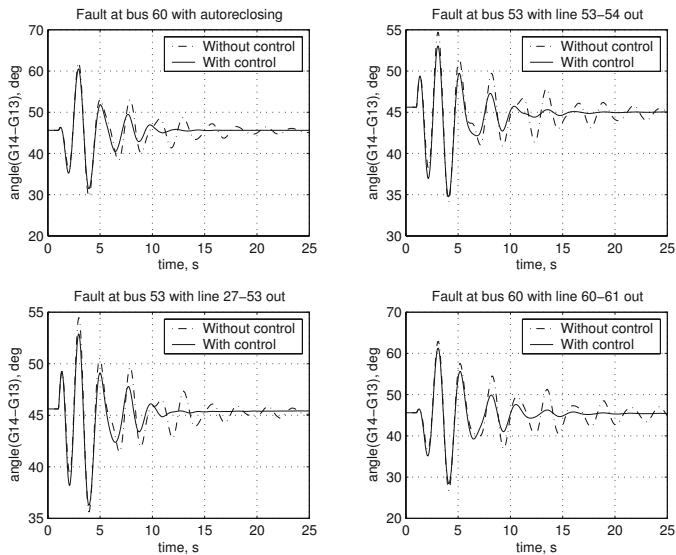


Fig. 13.9. Dynamic response of the system

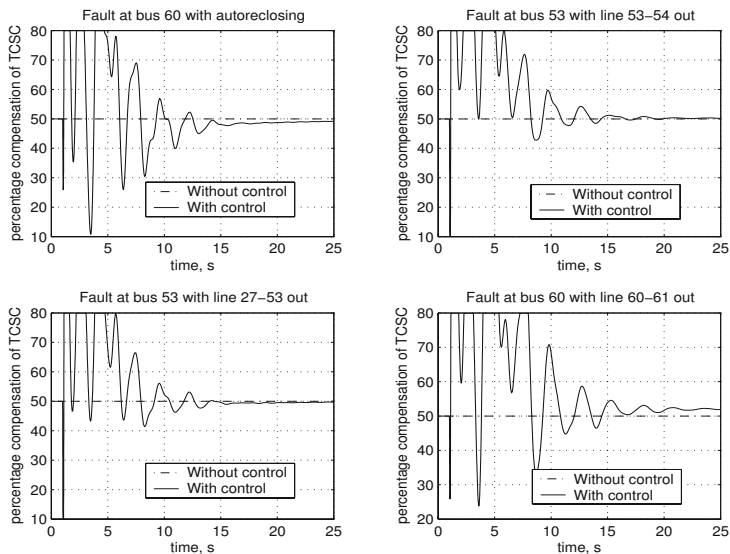


Fig. 13.10. Dynamic response of the system

13.6 Case Study on Sequential Design

In this section, a case study considering sequential design of damping controllers for multiple FACTS-devices is presented. The basic control design formulation is exactly the same as in the previous section. However, a separate controller is designed for each of the FACTS-devices sequentially. The feedback signals are chosen appropriately out of those locally available.

13.6.1 Test System

The study system described in chapter 12 is used again. Three FACTS-devices are considered to be installed as shown in Fig. 13.11. The TCSC is installed in the line between buses #18 and #50 to provide compensation (k_c) of 50%. An SVC is present at bus #18 to provide voltage support in the presence of the 1500 MW power transfer between area #5 and NYPS. The SVC is set to provide 117 MVar to ensure nominal voltage at bus #18. A TCPS with a steady state phase angle (ϕ) setting of 10 degrees is installed in the line connecting buses #13 and #17 to facilitate 3000 MW power transfer from the equivalent generation G13 to the rest of the NYPS. The aim of this exercise is to design three separate damping controllers K_1 , K_2 and K_3 , which use locally available signals only, such that inter-area oscillations are damped. The location of the FACTS-devices and the corresponding damping controllers are shown in Fig. 13.11, where y_1 , y_2 and y_3 are the measured feedback signals and u_1 , u_2 and u_3 are the derived control signals.

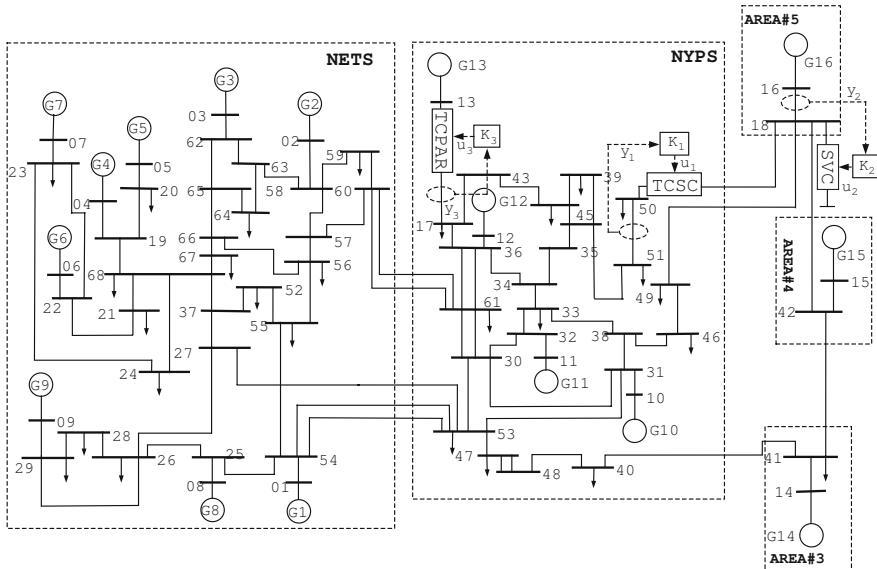


Fig. 13.11. Sixteen machine five area study system with three FACTS-devices

13.7.2 Control Design

The control design formulation described in Section 13.6.2 produces centralized controllers in multi-variable form. The design is now to be carried out in a sequential manner. The basic idea is to design a damping controller for one device to start with. The closed-loop system using this controller is used to design the controller for the second device. Exactly the same procedure is repeated for the third device. At each stage of this sequential design, the system model is updated with the designed controller model. In this process, the order of the system increases as each loop is closed depending on the number of states associated with the controllers of the individual FACTS-devices.

The sequential design of the controllers K_1 , K_2 and K_3 for the TCSC, SVC and TCPS has been carried out in sequence. The choice of this sequence improves the damping of modes #1, #2 and #3 in that order. Other sequences were tested and found to produce slightly different controllers but, in each case, similar performance was achieved. The same set of weights given in (13.29) and (13.30) has been found to work well for the design of all three controllers.

$$W_1(s) = 0.8475 \frac{99s + 11400}{s^2 + 156s + 12504} \quad (13.29)$$

$$W_2(s) = 0.8475 \frac{0.1055s^2 + 0.037s + 0.0094}{s(s + 0.0020)^2} \quad (13.30)$$

Each of the controllers was reduced to lower order by balanced truncation without significantly affecting the frequency response. The gains of the controllers (but not the controller structure) were scaled slightly to produce a damping ratio that ensured the oscillations settled in 10-12 seconds.

13.6.3 Performance evaluation

The eigenvalues of the closed-loop system produced by sequential loop closure were examined. Table 13.4 shows the eigenvalues considering only the controller K_1 for the TCSC. The damping of mode #1 shown in boldface, is improved primarily with very little effect on modes #2, #3 and #4. Similarly Table 13.5 shows that the controller primarily associated with the SVC improves the damping of mode #2, shown in boldface, besides improving mode #1 slightly. Finally, as shown in Table 13.6, the controller for the TCPS mainly improves the damping of mode #3, shown in boldface, besides adding to the damping ratios of modes #1 and #2. The combined action of the three controllers improves the damping of all three critical inter-area modes to adequate level.

Table 13.4. Damping ratios and frequencies of inter-area modes with the controller of the TCSC (Control loops of the SVC and TCPS open)

Mode No	Open-loop		Closed-loop	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
1	0.0626	0.3945	0.1544	0.3434
2	0.0434	0.5105	0.0545	0.4991
3	0.0560	0.6269	0.0656	0.6191
4	0.0499	0.7923	0.0502	0.7918

Table 13.5. Damping ratios and frequencies of inter-area modes with the controllers of the TCSC and SVC (control loop of the TCPS open)

Mode No	Open-loop		Closed-loop	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
1	0.1544	0.3434	0.1795	0.3158
2	0.0545	0.4991	0.1031	0.4549
3	0.0656	0.6191	0.0643	0.6184
4	0.0502	0.7918	0.0603	0.7864

Table 13.6. Damping ratios and frequencies of inter-area modes with the controllers of the TCSC, SVC and TCPS (all the control loops closed)

Mode No	Open-loop		Closed-loop	
	ξ	$f(\text{Hz})$	ξ	$f(\text{Hz})$
1	0.1795	0.3158	0.3140	0.2682
2	0.1031	0.4549	0.2266	0.4444
3	0.0643	0.6184	0.1105	0.4585
4	0.0603	0.7864	0.0600	0.7858

13.6.4 Simulation Results

As a part of the performance testing and validation exercise, a non-linear simulation was carried out under the same set of operating conditions described in section 13.5.4. Again, a series of disturbances, consisting of a solid three-phase fault lasting for 80 ms (5 cycles) followed by the contingency conditions depicted in Figures 13.8 to 13.10, were applied to the system.

The angular separation between machines G1 and G15 located in different areas is shown in Fig. 13.12 under different operating scenarios. In each case, the designed controllers of the TCSC, SVC and TCPS are able to settle the oscillations within 12-15 s. The outputs of the individual FACTS-devices are shown in Fig. 13.13, 13.14 and 13.15 for the same operating conditions. Appropriate limits were imposed on the variation of the control variables, as seen from their output response. The limit imposed on the TCSC is the same as before. For the SVC, the output variation limit was set to -150 (inductive) to 200 (capacitive) MVar. The limit on the phase angle of the TCPS was set to between 0 and 20 degrees.

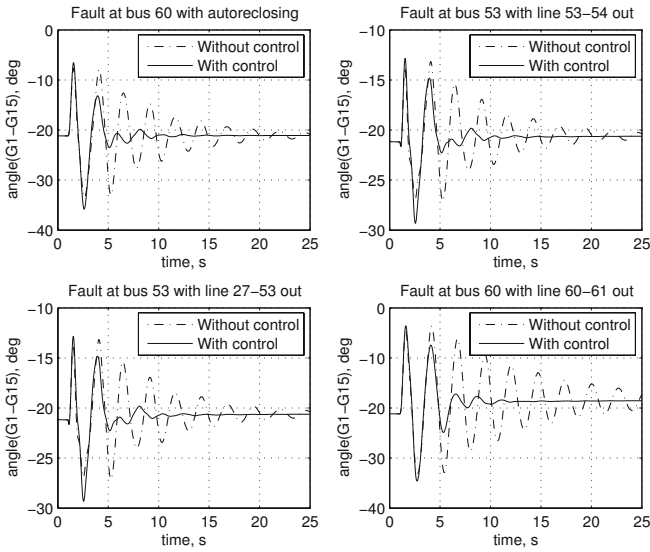


Fig. 13.12. Dynamic response of the system: angle between G1 and G15

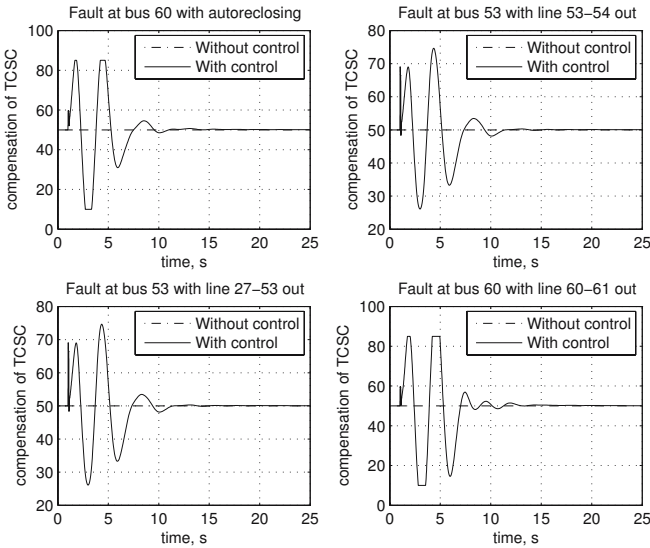


Fig. 13.13. Percentage compensation of the TCSC

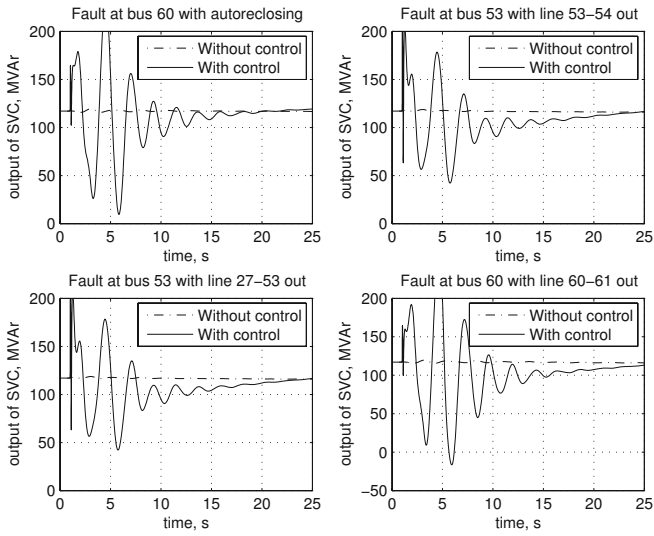


Fig. 13.14. Output of the SVC

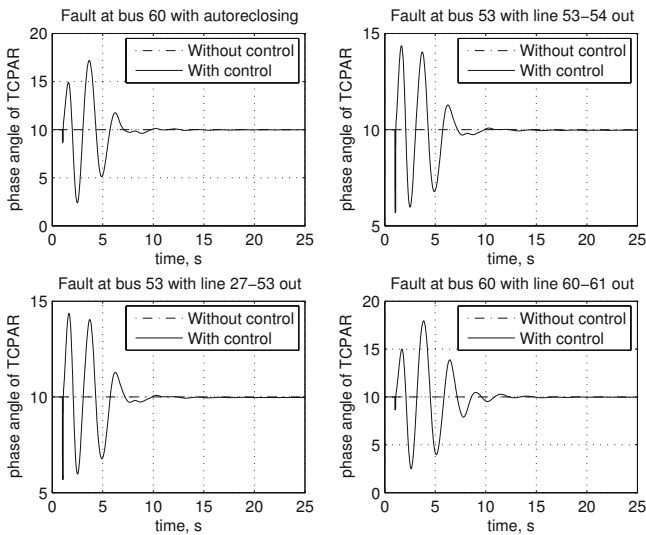


Fig. 13.15. Phase angle of the TCPS

13.7 H-Infinity Control for Time Delayed Systems

In large scale power systems, inter-area response can be damped more effectively through the use of wide-area measurement systems (WAMS), like introduced in chapter 11. Optical fiber communication is the right kind of technology for sensing and measurement systems used to implement the wide-area measurement systems. The advent of global positioning system (GPS) technology has made time stamping fairly routine and measurements of phase and other temporal information are, therefore, attainable through the use of commercially available equipment [24]. The information architecture, proposed in [25] is capable of providing timely, secure, reliable information exchange among various entities in a power system.

With the rapid advancements in Wide Area Measurement System (WAMS-) Technologies coupled with a fast and reliable data transmission infrastructure, the prospect of centralized control of power systems has gained momentum. Damping of inter-area oscillations through remote measurements, therefore, is becoming more feasible. From an economic viewpoint, implementation of centralized control using remote signals often turns out to be more cost effective than installing new control devices [26]. The obvious question is, at what speed are these remote measurements available to the control site? Employing phasor measurement units (PMUs), it is possible to deliver the signals at a speed as high as a 30 Hz sampling rate [3],[24]. It is possible to deploy the PMUs at strategic locations of the grid and obtain a coherent picture of the entire network in real time [24]. Many utilities in the USA and Europe have already installed demonstration projects.

The infrastructure cost and associated complexities, however, restrict the use of such sophisticated signal transmission hardware on a larger commercial scale. As a more viable alternative, the existing communication channels can be effectively used to transmit the signals from remote locations. The major problem is the delay involved between the instant of measurement and that of the signal being available to the controller. In the previous sections, no time-delay was considered as fast transmission of the necessary signals (typically within 0.02-0.05 s) was assumed. As the delay was considerably less than the smallest time period of the inter-area modes, it was not necessary to consider it in the design stage. A conservative estimate of the delay can typically be in the range of 0.5-1.0 s depending on the distance, transmission protocol and several other factors. Such a long delay should be accounted for in the design stage itself, to ensure effective control action.

The power system is treated as a dead-time system involving a long delay in transmitting the measured signals from remote locations to the controller site. It is not straightforward to control such time-delayed systems [27]. Normal H_∞ controllers are unlikely to guarantee satisfactory control of time-delayed systems. The Smith predictor [28][29] approach, proposed in the early fifties, was the first effective tool for handling such control problems. The difficulties associated with the design of H_∞ controllers for time-delayed systems and potential solutions using the predictor based approach is discussed later.

Since its arrival in the 1950s, a number of variations of the Smith predictor has been proposed in the literature. One of the drawbacks of the classical Smith predictor (CSP) approach is that it is very difficult to ensure a minimum damping ratio of the closed loop poles, when the open-loop system model has lightly damped poles (as often encountered in power systems). A modified Smith predictor (MSP) approach proposed in [30], was used to overcome the drawbacks of the CSP. However, the control design using the MSP approach might run into numerical problems for systems with fast stable poles. To overcome the drawbacks of CSP and MSP, a unified Smith predictor (USP) approach was proposed very recently by Zhong [31]. The USP approach effectively combines the advantageous features of both the CSP and MSP.

In this section, we illustrate the application of the USP approach for designing a damping controller for a power system with time-delayed signals. The predictor based control design methodology is illustrated by applying a case study to the same power system model described earlier. The performance and robustness of the designed controller is validated using frequency domain analysis and time domain simulations.

13.8 Smith Predictor for Time-Delayed Systems

In a time-delayed or dead-time system, either the measured output takes a certain time before it affects the control input or the action of the control input takes a certain time before it influences the measured outputs. Typical dead time systems consist of input and/or output delays. The general control setup for a system having an output delay is shown in Fig. 13.16, where,

$$P(s) = \begin{bmatrix} P_{11}(s) & P_{12}(s) \\ P_{21}(s) & P_{22}(s) \end{bmatrix} \quad (13.31)$$

The closed-loop transfer matrix from d to z is: $T_{zd} = P_{11} + P_{12}Ke^{-sh}(I - P_{22}Ke^{-sh})^{-1}P_{21}$. An equivalent structure is shown in Fig. 13.17. This suggests that there exists an instantaneous response through the path P_{11} (path 1 in Fig. 13.17) without any delay.

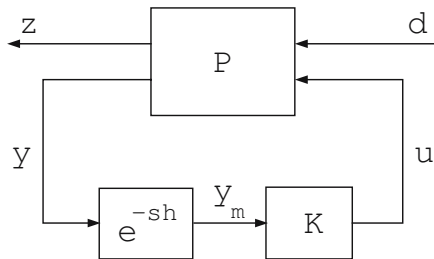


Fig. 13.16. Control setup for dead-time systems

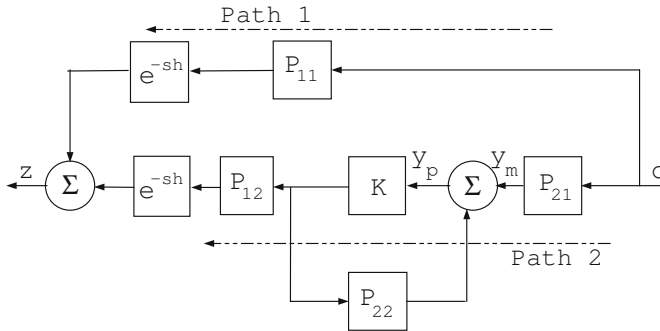


Fig. 13.19. Uniform delay in both paths

The second task of bringing a delay in path 1 is done while forming the generalized regulator prior to control design. The presence of the predictor block Z and the delay in path 1 ensures that the responses (through path 1 and path 2) governing the performance index are delayed uniformly, as shown in Fig. 13.19.

A predictor-based controller for the dead-time system $P_h(s) = P_{22}(s)e^{-sh}$ consists of a predictor $Z = P_{22}(s) - P_{22}(s)e^{-sh}$ and a stabilizing compensator K , as shown in Fig. 13.20. The predictor Z is an exponentially stable system such that $P_h + Z$ is rational i.e. it does not involve any uncontrollable response governing the H_∞ performance index.

The shortcomings of the CSP approach for systems having poorly damped open-loop poles are overcome with the MSP approach [27]. Let us consider a generalized delay-free system given by equation 13.32.

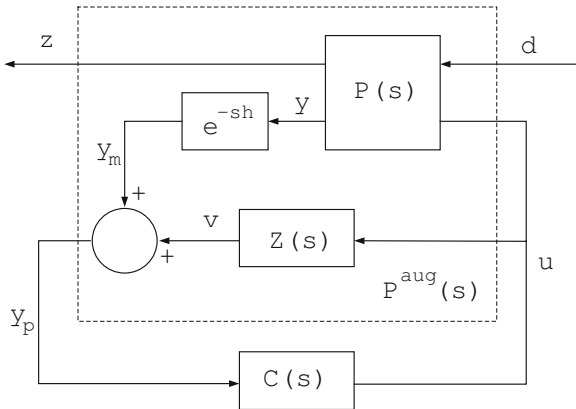


Fig. 13.20. Smith predictor formulation

$$P(s) = \begin{bmatrix} A & B_1 & B_2 \\ C_1 & D_{11} & D_{12} \\ C_2 & D_{21} & D_{22} \end{bmatrix} = \begin{bmatrix} P_{11} & P_{12} \\ P_{21} & P_{22} \end{bmatrix} \quad (13.32)$$

For a delay of h seconds, the generalized regulator formulation using the MSP approach [27] is as follows:

$$\tilde{P}(s) = \begin{bmatrix} A & e^{Ah}B_1 & B_2 \\ C_1 & 0 & D_{12} \\ C_2e^{-Ah} & D_{21} & 0 \end{bmatrix} \quad (13.33)$$

where $\tilde{P}(s)$ is $P_{aug}(s)$ including the effect of the delay block between d and z , as shown in Fig. 13.20.

The computation of matrix exponential e^{Ah} in (13.33) suffers from numerical problems, especially for systems having fast stable eigen-values. In the worst case it might well be non-computable. This problem can even arise with reasonably small amount of delay, if some of the stable eigenvalues are very fast.

In H_∞ mixed-sensitivity formulation for power system damping control design, the presence of fast stable eigenvalues in the augmented system cannot be ruled out. Possible sources of fast stable eigenvalues include the fast sensing circuits ($T \sim 0.02s$), fast damper circuits ($T \sim 0.05s$) and even the weighting filters. These often lead to numerical instability when solving the problem using LMIs. These problems are overcome through the use of the USP [31] formulation, achieved by decomposing the delay free system model P into a critical part P_c and a non-critical part P_{nc} . The critical part contains the poorly damped poles of the system, whereas the non-critical part consists of poles with sufficiently large negative real values. The next section describes the generalized problem formulation using this approach.

13.9 Problem Formulation using Unified Smith Predictor

As indicated in the previous section, the first step towards formulating the control problem using the USP approach is to decompose the delay-free system model into critical and non-critical parts. This is normally done by applying a suitable linear coordinate transformation on the state space representation of the system. In this work, a suitable transformation matrix V is chosen such that the transformed matrix $J = V^{-1}AV$ is in the Jordan canonical form and is free from complex entries. The transformation matrix V is chosen using the *eig* function available in Matlab [18]. The elements of the transformed matrix J were converted from complex diagonal form to a real diagonal form using the *cdf2rdf* function in Matlab [18]. The transformed augmented delay-free system model P_{22}^T is given by:

$$P_{22}'(s) = \begin{bmatrix} V^{-1}AV & V^{-1}B_2 \\ C_2V & D_{22} \end{bmatrix} = \begin{bmatrix} A_c & 0 & B_c \\ 0 & A_{nc} & B_{nc} \\ C_c & C_{nc} & D_{22} \end{bmatrix} \quad (13.34)$$

where A_c is the critical and A_{nc} is the non-critical part of A . The augmented system model P_{22}' can be split as $P_{22}' = P_c + P_{nc}$, where:

$$P_c(s) = \begin{bmatrix} A_c & B_c \\ C_c & D_{22} \end{bmatrix} \quad (13.35)$$

$$P_{nc}(s) = \begin{bmatrix} A_{nc} & B_{nc} \\ C_{nc} & 0 \end{bmatrix} \quad (13.36)$$

The predictor for the critical part is formulated using the MSP approach by applying a completion operator [27]. On a rational transfer matrix $G = C(sI - A)^{-1}B + D$, the completion operator $\pi_h\{e^{-sh}G\}$ is defined as follows:

$$\pi_h\{e^{-sh}G\} = \begin{bmatrix} A & B \\ C_e e^{-Ah} & 0 \end{bmatrix} - \begin{bmatrix} A & B \\ C & D \end{bmatrix} e^{-sh} \quad (13.37)$$

Using (13.37), the predictor for the critical part P_c is given by (13.38). More details are provided in reference [27].

$$Z_c(s) = \pi_h\{e^{-sh}P_c\} = \begin{bmatrix} A_c & B_c \\ C_c e^{-A_c h} & 0 \end{bmatrix} - \begin{bmatrix} A_c & B_c \\ C_c & D_c \end{bmatrix} e^{-sh} = P_c^{\text{aug}}(s) - P_c(s)e^{-sh} \quad (13.38)$$

The predictor for the non-critical part is constructed following the CSP formulation and is given by:

$$Z_{nc}(s) = P_{nc}(s) - P_{nc}(s)e^{-sh} \quad (13.39)$$

The USP, denoted by Z , is simply the sum of Z_c and Z_{nc} , as shown in Fig. 13.21.

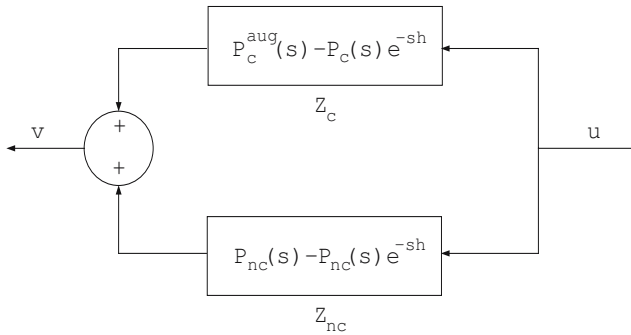


Fig. 13.21. Unified Smith predictor

It is given by:

$$Z(s) = P_{22}^{aug}(s) - P_{22}^t(s)e^{-sh} \quad (13.40)$$

where $P_{22}^{aug} = P_{nc} + P_c^{aug}$. Using (13.34), (13.36) and (13.38), the realization for P_{22}^{aug} can be expressed in the form:

$$P_{22}^{aug} = \begin{bmatrix} A & B_2 \\ C_2 E_h & 0 \end{bmatrix} \quad (13.41)$$

where

$$E_h = V \begin{bmatrix} e^{-A_c h} & 0 \\ 0 & I_{nc} \end{bmatrix} V^{-1} \quad (13.42)$$

The augmented system model P^{aug} is obtained by connecting the original dead-time system and the USP in parallel, as shown in Fig.13.21. The new set of measured outputs is y_p . The expression for P^{aug} is given by:

$$P^{aug} = \begin{bmatrix} P_{11}(s) & P_{12}(s)e^{-sh} \\ P_{21}(s) & P_{22}(s)^{aug} \end{bmatrix} \quad (13.43)$$

The generalized regulator $\tilde{P}$ can be formulated from P^{aug} after inserting the delay block e^{-sh} in between d and z , as shown by a dotted box in Fig. 13.21. The steps to arriving at the final expression for $\tilde{P}$ are detailed in [31]. The final form of the generalized regulator is as follows:

$$P = \begin{bmatrix} A & 0 & E_h^{-1} B_1 & B_2 \\ 0 & A_{nc} & [0 \quad e^{A_{nc} h} - I_{nc}] V^{-1} B_1 & 0 \\ C_1 & C_1 V \begin{bmatrix} 0 \\ I_{nc} \end{bmatrix} & 0 & D_{12} \\ C_2 E_h & 0 & D_{21} & 0 \end{bmatrix} \quad (13.44)$$

Having formulated the generalized regulator $\tilde{P}$, following the USP approach, the objective is to design a controller K to meet the desired performance specifications. If K ensures the desired performance for $\tilde{P}$ then the controller predictor combination $K_e = K(I-ZK)^{-1}$ is guaranteed to achieve the same for the original dead-time system [27].

13.10 Case Study

In this section, the prototype power system model described earlier is considered to illustrate the control design methodology in details. The performance and ro-

bustness of the design is also validated. A conservative estimate of 0.75 s was considered as the signal transmission delay, with a view to cover the worst case scenario. We have used an SVC as the FACTS-device in the system.

13.10.1 Control Design

The control design problem was formulated using the standard mixed-sensitivity approach [16][17] with modifications to include the effect of delay. The overall control setup is shown in Fig. 13.22, where:

- $G_p(s)$: power system model,
- $G_{svc}(s)$: SVC model,
- $W_1(s), W_2(s)$: weighting filters,
- $K(s)$: controller to be designed,
- $Z(s)$: Smith predictor,
- d : disturbance at the system output,
- z : weighted exogenous outputs,
- y_m : measured output and
- u : control input.

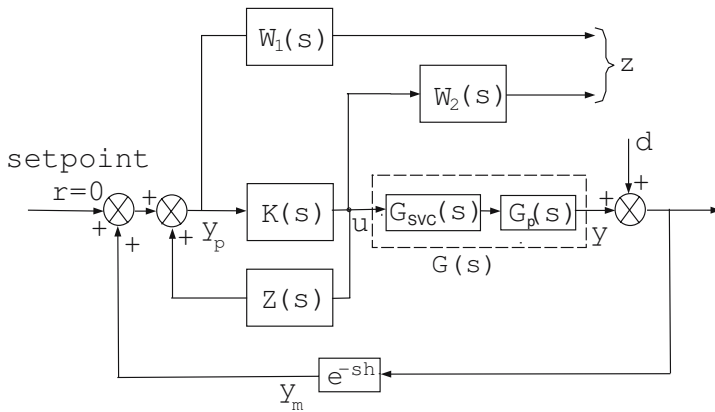


Fig. 13.22. Control setup with mixed-sensitivity design formulation

The design objective is as follows:

Find a controller K from the set of internally stabilizing controllers S such that:

$$\min_{K \in S} \left\| \begin{bmatrix} W_1 S \\ W_2 K S \end{bmatrix} \right\|_{\infty} < 1 \quad (13.45)$$

where $S = (I - GK)^{-1}$ is the sensitivity. The solution to the problem was sought numerically using the LMI solver with an additional pole-placement constraint.

Once the generalized regulator is formulated following the steps given in section 13.3, the basic steps for control design are exactly similar to those outlined earlier. The weights were chosen as follows:

$$W_1(s) = \frac{100}{s + 100}, W_2(s) = \frac{100s}{s + 100} \quad (13.46)$$

The frequency responses for the weighting filters are shown in Fig. 13.23. They are in accordance with the basic requirement of mixed-sensitivity design.

To facilitate the control design and to reduce the complexity of the designed controller, the nominal system model was reduced to a lower order. Using the simplified system model and the above-mentioned weights, the generalized problem was formulated according to (13.44). The solution was sought numerically using suitably defined objectives in the argument of the function *hinfnmix* of the *LMI Toolbox* in Matlab [18]. The pole-placement constraint was imposed by using a 'conic sector' of inner angle $2\cos^{-1}(0.175)$ with apex at the origin. The order of the controller obtained from the LMI solution was equal to the reduced order augmented system order plus the order of the weights. The designed controller was reduced to an 8th order equivalent using Schur's method without affecting the frequency response as shown in Fig. 13.24.

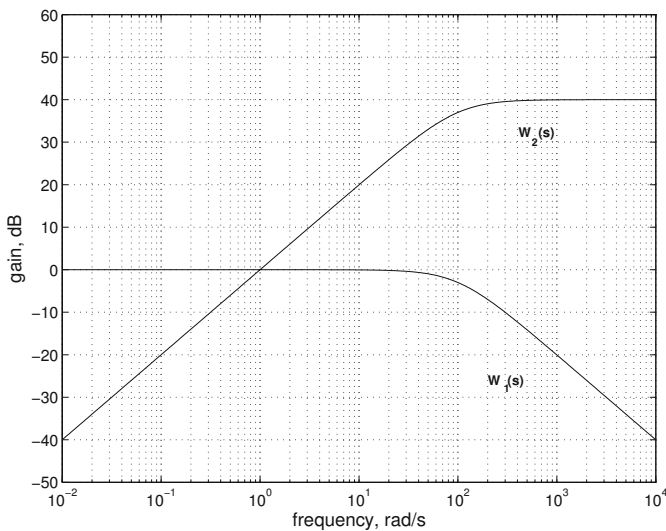


Fig. 13.23. Frequency response of the weighting filters

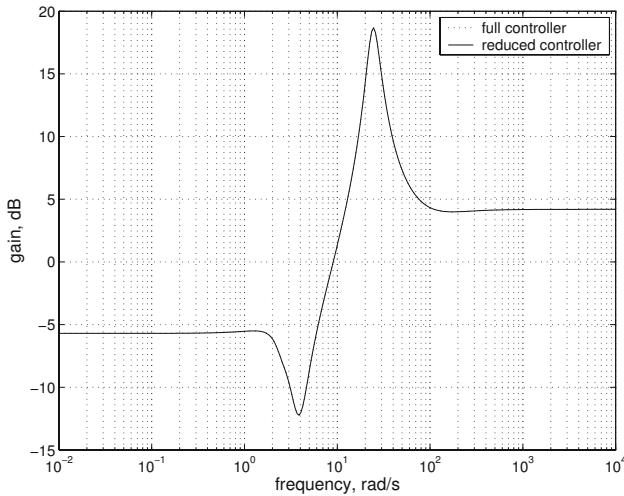


Fig. 13.24. Frequency response of the full and reduced controller

13.10.2 Performance Evaluation

The eigenvalues of the closed-loop system were computed to examine the performance of the designed controller in terms of improving the damping ratios of the inter-area modes. A 4th order Pade approximation was used to represent the delay ($= 0.75$ s) in the frequency domain.

13.10.3 Simulation Results

The controller for the SVC was designed following exactly the same procedure as used for the TCSC earlier in this chapter. The same disturbance, as considered in the previous section, was examined. The dynamic response of the system following this disturbance is shown in Fig. 13.25. This figure illustrates the relative angular separation between the generators G1 and G15. It is clear that the inter-area oscillation is damped out in 12–15 s, even though the feedback signals arrive at the control location after a finite time delay of 0.75 s. The variation of the output of the SVC is shown in Fig. 13.26. It is within a range of -1.5 p.u. to 2.0 p.u.

The performance of the controller for different delays is shown in Figs. 13.27 and 13.28. The simulation results in Fig. 13.29 demonstrate the detrimental effect of not considering the delay at the design stage, if there is one in practice. The results highlight a potential application of the USP approach for power system. It could be used for the design of damping control with different types of FACTS-devices, where the transmission of remote feedback signals involves a finite amount of time delay.

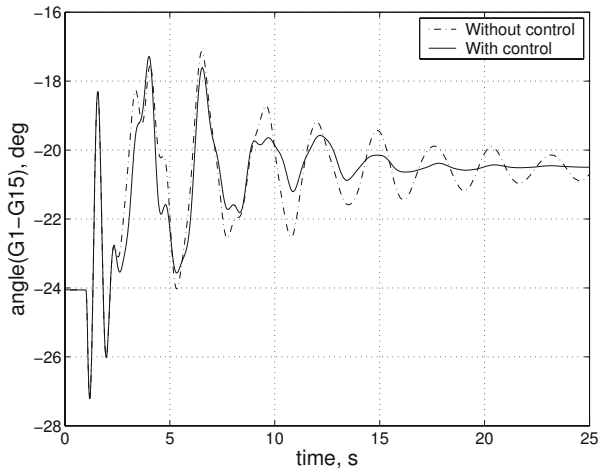


Fig. 13.25. Dynamic response of the system with SVC; controller designed considering delay

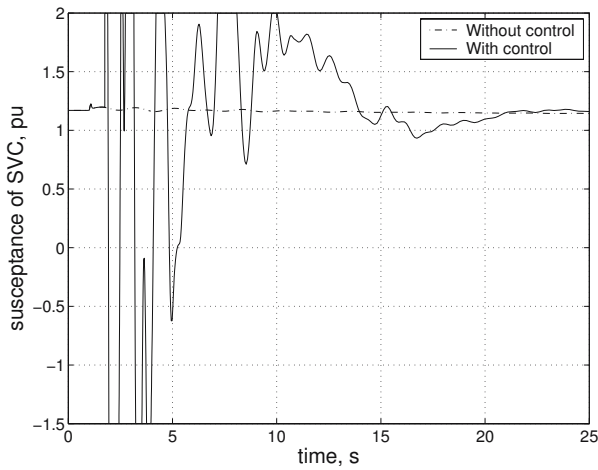


Fig. 13.26. Output of the SVC

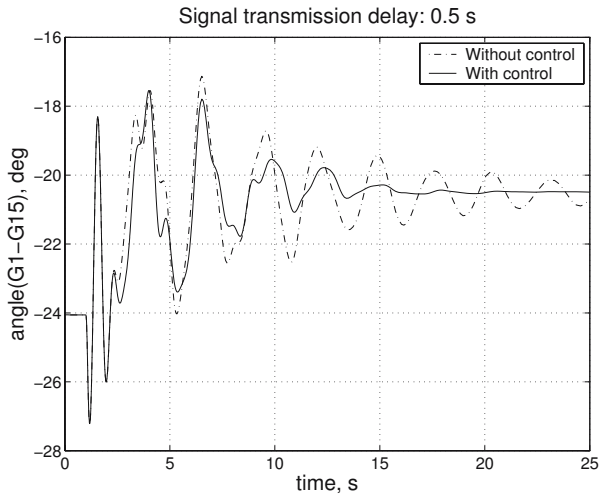


Fig. 13.27. Dynamic response of the system with a delay of 0.5 s

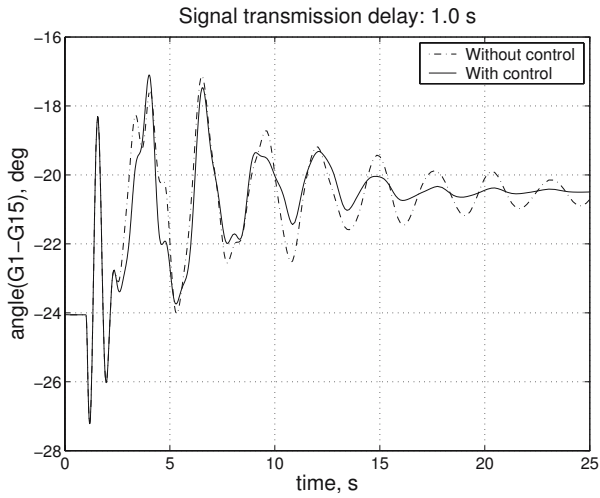


Fig. 13.28. Dynamic response of the system with a delay of 1.0 s

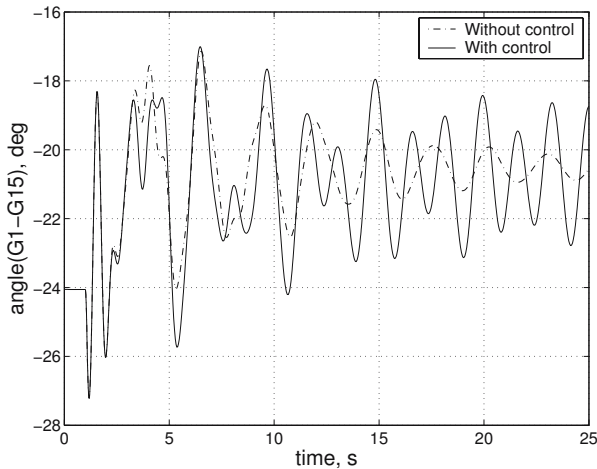


Fig. 13.29. Dynamic response of the system; controller designed without considering delay

In this chapter, the basic concept of mixed-sensitivity design formulation has been elaborated. The problem has been translated into a generalized H_∞ problem. The solution to the problem was sought numerically using LMIs with pole-placement. The control design methodology was illustrated by two case studies. In the first case, a centralized controller was designed for a single FACTS-device. Feedback signals from three remote locations were employed for the controller. In the second case, a sequential design methodology was adopted for multiple FACTS-devices. Local feedback signals were used to design decentralized controllers for individual FACTS-devices. The performance and robustness of the design was validated using frequency domain analysis and non-linear simulations.

A methodology for power system damping control design that accounts for delayed arrival of feedback signals from remote locations is also described. A predictor based H_∞ control design strategy has been presented for such time-delayed systems. The design procedure based on the USP approach has been applied for the centralized design of a power system damping controller for FACTS-devices such as TCSC and SVC. A combination of the USP and the designed controller was found to work satisfactorily under different operating scenarios, even though the stabilizing signals could reach the controller site only after a finite time.

In this chapter, a fixed time delay has been considered for all the communication channels. In practice, this might not always be the case as the distances from the measurement sites differ. Therefore, different amount of delay for each signal needs to be considered during the design. Further research is currently being conducted in this area.

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